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(54) High frequency pulse transformer for an IGBT gate drive

(57) A pulse transformer for use in a high frequency motor controller to isolate the electronic control stages from the power stages in a motor controller is claimed. The primary (40) and secondary (42) windings are wound on a common ferrite core (66) and spaced apart from one another. The wire size used for the windings is relatively small (gauge 37 or less). Each of the windings has between 10 and 13 turns. The inter-winding capacitance must be below 5 picofarads, preferably below 1 picofarad, and most preferably below 0.2 picofarad.

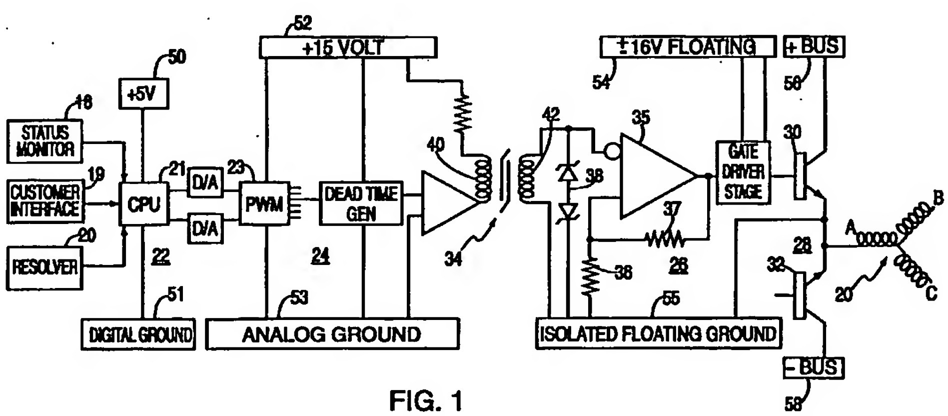

FIG. 1

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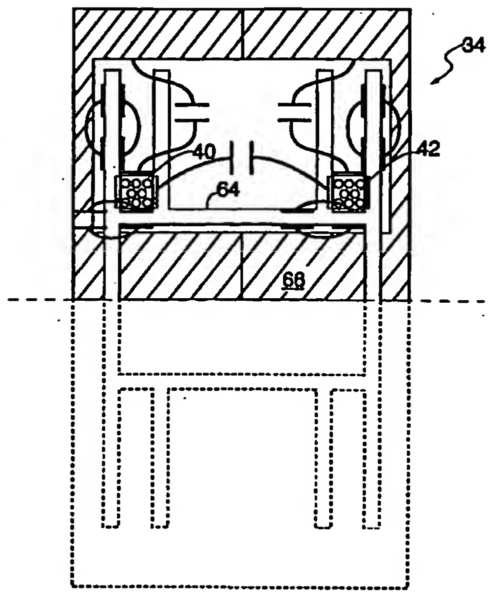


FIG. 2

Description

This invention relates to a high frequency pulse transformer, and more particularly, to a high frequency pulse transformer for an IGBT gate drive in a motor controller.

BACKGROUND OF THE INVENTION

US-A-4,686,347 describes a motor control system whereby a microprocessor or similar electronics can be used to control electric motors. The sensitive electronics are coupled to the motor via a power transistor switching stage wherein the transistors operate in on/off fashion to control the current flow to the motor windings in accord with control signals from the microprocessor. The microprocessor receives feedback from the motor indicating rotor position and other controllable parameters. The microprocessor may also receive command signals usable for servo control. The command signals and the feedback signals are compared to determine the frequency, amplitude and phase for the winding energization, and the rotor position feedback is used to determine the proper commutation of the winding excitation. The microprocessor, either alone or with other components, provides on/off control signals to the power stages which, in turn, supply current to the windings with the desired amplitude, frequency and phase commutated according to the rotor position. Most present day servo control systems utilize this basic control system.

US-A-4,447,771 and US-A-4,490,661 disclose motor control systems in which digital electronics control power switching transistors via a pulse width modulator (PWM) so that the excitation current to the motor is a function of the pulse width. Both patents disclose techniques for controlling the phase angle of the excitation current relative to the rotor position to improve the torque efficiency of the system.

Most present day systems use pulse width modulation as disclosed in US-A-4,447,771 and US-A-4,490,661 to convert the control signals from the sensitive electronics to the desired excitation currents for the motor windings. In such systems used to control large motors isolation is necessary between the sensitive electronic controller and the brute force power stages that drive the motor. Optical coupling is used to provide a high degree of isolation in most present day controller designs which operate at a PWM switching rate of about 2 KHz.

There is an ever present desire to provide smaller and less costly controllers and controllers with high overall system efficiency. The invention provides those advantages for any size motor. However, the advantages are more pronounced for large motors in the range of 10 to 150 horsepowers and are most pronounced for the very large motors in the 150 to 500 horsepower range. In the higher horsepower ranges, inefficient use of components and inefficient overall system design directly translate into high controller and high usage costs.

One approach to smaller and less costly controllers is to increase the switching frequency of the PWM power stage. In theory at least, higher switching frequencies should make it possible to control the needed power for large motors using smaller an less expensive components. However, the problems associated with operating the power stages at higher switching frequencies have proven to be either insolvable or prohibitively costly both in space and money. High frequency noise can bleed off power from the system and seriously affect the system efficiency. Noise of the common mode variety noise that goes up and down together on a parallel conductor pair) can be particularly difficult to detect and even more difficult to eliminate.

SUMMARY OF THE INVENTION

It is an object of the invention to provide improved isolation between the sensitive electronic stages and the associated power stages.

It is another object of the invention to provide effective isolation coupling between the electronic stages and the associated power stages in a motor control system where the power stage can operate at a switching rate in the range of above 6000 Hz.

Still another object of the invention is to provide an isolation transformer for a motor controller operating at a high power switching rate in the range of above 6 KHz with little high frequency power loss and negligible noise generation.

Yet another object is to provide a pulse transformer for a motor controller with higher system efficiency, i.e., maximum torque, relative to energy applied to the controller.

These and other objects are achieved by the high frequency pulse transformer as claimed in Claims 1 to 8.

Optical coupling which provides effective coupling isolation at low switching rates, for example, about 2000 Hz, has too much time delay for effective operation in the range above 6000 Hz. Furthermore, switching at the 6000 Hz rate can generate transients as high as 100 MHz which are not effectively decoupled with the optical devices. Known high frequency, electronic transformer coupling is too expensive and complicated for effective use in a motor controller.

The pulse transformer in the form of a dynamic coupling inductance according to the invention includes spaced apart primary and secondary windings wound around a common high permeability core. The flux return path for the flux through the core is configured to surround the windings. The core is configured so that it is saturable in normal operation. The windings are spaced as far apart as possible to minimize the interwinding capacitance. The windings are wound using a small number of turns and small wires to minimize the surface area of the winding to thereby reduce winding to

shell and winding to winding capacitance.

If a rectangular pulse is applied to the primary winding of the pulse transformer, a positive spike appears at the secondary winding corresponding to the rise time and a negative spike appears according to the fall time. The rise time of the positive spike occurs while the core is being driven into saturation and the windings are coupled to one another. Once saturation is reached the windings are decoupled. A similar operation takes place during the fall time. By careful design according to the concept of the invention, turn on and turn off times for pulses passing through the pulse transformer are as little as 100 nanoseconds which is about 10 times faster than could be accomplished using optical coupling.

Rectangular pulses from a microprocessor and other electronic stages pass through the saturable pulse transformer as a positive and a negative spike pair which is supplied to a Schmitt trigger that reconstitutes the rectangular pulse for control of the power transistors. With the arrangement according to the invention, only the leading and trailing edges of the pulses pass through the pulse transformer. As a result, power stage switching can be achieved at the rate of 10A/100 nanoseconds. Using conventional components, switching for large motor control systems can be on the order of about 250 to about 450 nanoseconds, depending on the motor size. Faster switching speeds, as fast as 150 nanoseconds, can be achieved by using special high speed components.

The coupling between the electronic and power stages according to the Invention results in a very high rejection of common mode noise and elimination of the related energy absorption. Contrary to conventional design, the primary and secondary windings are located on a ferrite core and separated as far as possible to minimize inter-winding capacitance. A minimum number of winding turns and a small diameter wire are used for the winding to minimize the conductive surface area. Although this structure provides relatively poor magnetic coupling between the windings, it is sufficient since only the leading and trailing edges of the pulses actually pass through the transformer.

BRIEF DESCRIPTION OF THE DRAWINGS

The foregoing and other objects are achieved with embodiments as described in the following specification which includes the drawings as follows:

Fig. 1 is a block diagram showing the overall layout of a motor controller according to the invention.

Fig. 2 is a cross-sectional view of the high frequency pulse transformer according to the invention.

Fig. 3 is a schematic diagram of the electronic stages connected to the primary side of the pulse transformer.

Fig. 4 is a schematic diagram of the gate drive including the pulse transformer and Schmitt trigger circuit connected to the secondary side of the pulse transformer.

Fig. 5 is a schematic diagram of the power supplies for the motor controller.

Fig. 6 is a schematic diagram of the independent floating ground supplies for the gate drives.

DETAILED DESCRIPTION

The motor controller illustrated in Fig. 1 controls a brushless motor **20** having a three phase stator winding including winding phases A, B and C. The winding phases are each displaced by 120 electrical degrees. Although the invention can be used with any type of brushless motor, it is particularly useful with a motor having three phase stator windings operating in the current pulse range above 2000 Hz and preferably in the range of 6000 to 20000 Hz or higher. Although the invention is useful with any size motor, the advantages are particularly important in controlling large and very large motors.

The motor controller includes a digital section **22** including a central processing unit **21** (CPU), such as a microprocessor, to receive control commands and motor feedback signals to generate sinusoidal signals corresponding to the frequency, phase and amplitude of the motor excitation currents. The signals from the digital section pass through an analog section **24** including a PWM **23** to convert the sinusoidal signals into corresponding pulse trains of pulses with varying width. The pulse trains pass through a gate drive **26** to the main power transistors **28**.

The digital section generates two sinusoidal signals for the three phase motor excitation, the third signal being derived in analog section **24** as the sum of the other two phases. The analog section generates complimentary pulse trains for each of the three phases, or a total of six pulse trains, only one of which is shown in Fig. 1. The six pulse trains each pass through a separate gate drive (only one being shown) to control the conductive state of the six power transistors (only two of which are shown).

The microprocessor can be programmed to produce a sinusoidal excitation current of any desired phase, amplitude and frequency to almost any type of motor including synchronous, induction or reluctance motors. The system can also be programmed to calculate excitation phase angle for maximum torque producing efficiency according to the aforementioned US-Patents.

In the operation of the controller it is essential that the power stages, including gate drives **26** and power transistor stages **28**, be isolated from the sensitive electronic stages **22** and **24**. For a conventional controller operating at a PWM rate of about 2000 Hz, the necessary isolation can be achieved with an optical coupler. But to reduce the size of components, and to achieve other advantages, a higher PWM pulse repetition rate in the range of 6000 to 20000 Hz is used.

To achieve the necessary isolation without prohibitive leakage losses and noise problems, a specially constructed ferrite core transformer **34** with extremely low inter-winding capacitance is used in combination with a Schmitt trigger located in gate drive **26**. When a PWM pulse is applied to the pulse transformer, the leading edge of the pulse drives the core into saturation. While being driven into saturation, primary winding **40** is coupled to secondary winding **42** and produces a sharp spike on the secondary winding. The trailing edge of the applied pulse likewise drives the core out of saturation and produces a sharp negative spike. Comparator **35** and resistors **36** and **37** are connected in a Schmitt trigger configuration. The positive spike turns the trigger on and the negative spike turns the trigger off. Thus, rectangular PWM pulses are passed to the gate drives by passing the leading and trailing edges through transformer **34** to the Schmitt trigger which reconstructs the rectangular pulses.

Not only it is necessary to isolate the power stages from the sensitive electronics stages, but it is also necessary to achieve similar isolation in the power supplies for the various stages. The CPU and other components of the digital section **22** require a +5 Volt supply **50** and ground return **51**. The PWM unit **23** and other components in analog section **24** require a +15 Volt supply **52** and ground return **53**. Each of the six gate drives **26** require a separate floating +16 Volt supply **54** with a floating ground **55**. The power for driving the motor is supplied to the power transistors from a +bus **56** and a -bus **58**. The structure of the various power supplies will be described hereinafter. However, the significant point is that the power stages connected to the output of isolation pulse transformer **34** have separate power supplies isolated from the power supplies for the electronic sections **22** and **24** connected to the primary of the pulse transformer **34**.

Pulse transformer **34** is shown in cross-sectional detail in Fig. 2. The transformer includes primary winding **40** and secondary winding **42** wound on a bobbin **64** so that the primary and secondary windings are well separated. A ferrite core **66** passes through the center of the windings and returns the outside of both windings. In other words, the core is constructed to completely surround the windings. Preferably, the core is a two piece structure, as shown in Fig. 2, with the core return path completely surrounding and enclosing the windings to provide self-shielding from radiated noise. The core material should be of a high permeability material to minimize the required number of winding turns. A powder ferric material type W provides good results. Materials such as Suppermeloy or Square 80 provide better results, but are too expensive for most applications.

The pulse transformer is designed to pass rectangular pulses with a repetition rate in the range of 6000 to 20000 Hz which include high frequency transients in the range of 1 MHz to 100 MHz. These high frequency transients can create severe power loss and noise problems. The winding configuration is critical to reducing the power loss and noise problems to acceptable levels. First, the primary and secondary windings are well separated and placed as far apart as possible to minimize the inter-winding capacitance (CWW). As shown in Fig. 2, a spacing of 200-300 mils, on the order of right times the width of the windings, is preferred. This spacing between the windings reduces the inter-winding capacitance (CWW). A higher wire gauge size (i.e., minimum diameter wire) such as gauge 37 is used to minimize winding surface areas which in turn minimize the winding to core capacitance (CWCC) and (CWCS). The number of winding turns is selected to produce an inductance of approximately 1 mH. Ten to thirteen turns have been found to produce the desired results. A larger number of turns increases the ratio relationship between the capacitance and inductance causing self oscillation which is not acceptable. Less than ten turns provide a ratio of capacitance to leakage inductance that attenuates the applied signal to an unacceptable degree. The total leakage capacitance according to the design of the invention has been measured at 0.08 to 0.2 picofarads which results in very low noise and very low high frequency power losses. To be effective, the pulse transformer must have a total leakage capacitance of 5 picofarads, and preferably below 1 picofarad. Conventional designs have been measured in the range of 32 to 43 picofarads and were found to be ineffective.

When a rectangular pulse is applied to the pulse transformer, the high permeability core quickly goes into saturation and produces a short positive spike in the secondary winding corresponding to the leading edge of the applied pulse. Similarly, the trailing edge of the applied pulse produces a sharp negative spike on the secondary winding. The rise time for the spikes is on the order of 100 nanoseconds. With these short time delays the pulse transformer according to the invention can be used to pass information for pulses in the range of 6000 to 20000 Hz with minimal radiation and noise losses.

Fig. 3 is a schematic diagram illustrating the electronic sections of the motor controller connected to supply the primary winding of pulse transformer **34**. These sections include the central processing unit (CPU) **21** and other digital components in digital section **22**. Also included is the pulse width modulator (PWM) **23** and other analog components in the analog electronics section **24**.

The central processing unit **21** can be a microprocessor or other type of digital processing electronics. The CPU receives rotor position information from the motor being controlled via a resolver or position encoder **17** which provides rotor position information. The rotor position information is converted to a digital form by a resolver-to-digital converter **20**. The position information is used to commutate the winding excitation calculated by the CPU according to rotor position. The position information can also be used for a servo loop positioning of the motor shaft. Also, the CPU can calculate motor speed from the rate of change of the rotor

position. The CPU receives data from a status monitor **18** which provides status information for any parameter being controlled and is useful in calculating the motor excitation. A customer interface **19** provides command information such as the desired speed, desired torque or desired position.

If the motor controller is operated as a velocity servo, for example, the CPU is programmed to compare the speed command with the feedback signal indicating actual speed to determine the torque required to achieve the desired speed. Likewise, servo position control can be achieved by comparing the desired position to the actual position to determine the motor torque required to reach the desired position. The CPU calculates an incrementally synthesized excitation signal including the frequency, phase and amplitude of the current excitation to be supplied to the motor. The CPU furnishes such a sinusoidal signal to a digital-to-analog (D/A) converter **70** for controlling the excitation at one phase of the motor. A second such signal, displaced by 120 electrical degrees, is supplied to another D/A converter **71** to control excitation for a second motor phase. The third phase is the sum of the other two phases and, therefore, need not be calculated by the CPU.

D/A converters **70** and **71** provide sinusoidal excitation signals to the PWM stage **23** which converts the excitation signals into pulse trains wherein the pulse width is proportional to the instantaneous signal amplitude. The pulse width determines the turn-on and turn-off times for the power transistor controlling current flow to the motor windings.

As described hereinafter, current sensors are provided for sensing the current flow to the motor windings to provide corresponding voltage feedback signals **77** and **81**. The feedback signal voltages are compared with the desired excitation signal values calculated by the CPU. The comparison of the calculated value for phase A and the actual feedback value of the current are compared and the result of the comparison appears as the output of amplifier **74**. Similarly, the feedback signal **81** is applied to the summing junction of amplifier **78** for comparison with the calculated signal applied to resistor **79** so that the results of the comparison appear at the output of amplifier **78**. The signals for phase A and C at the outputs of amplifiers **74** and **78**, respectively, are summed in operational amplifier **82** and summing resistors **83** and **84**. Since the amplifiers **74** and **78** produce signals displaced by 120 electrical degrees, the remaining phase B is derived by summing phases A and C in amplifier **82**.

The PWM pulse train is generated by comparing the excitation signal value from one of amplifiers **74**, **78** and **82** to a triangular waveform value generated by triangle generator **84**. The signal value comparisons are achieved by comparators **85**, **86** and **87** for phases A, B and C, respectively. More specifically, the excitation signal value for phase A is compared with triangle waveform generator **84** such that when the excitation signal value is below the triangular wave value, the output of comparator **85** is ON, and when the signal value is above the triangular wave value, the output of comparator **85** is OFF. The greater the excitation signal value, the wider the pulse emerging from comparator **85** and, therefore, the greater the current supplied to the motor winding. The smaller the signal value, the narrower the pulse and the smaller the current supplied to the motor winding. PWM pulse trains for phases B and C are generated in similar fashion.

The base frequency for the triangular wave in most current systems is on the order of 2000 Hz or less. In the motor controller according to the invention, the repetition rate of the triangular wave can be in the range of 6000 to 20000 Hz. The higher to frequency of the PWM pulses the smoother the control and the smaller the size of the components. Higher PWM frequencies can be employed using the same pulse transformers by using faster Schmitt triggers and gate drives.

To simplify the illustration, only the phase B section for the "dead" time generation unit **90** and the pulse transformer driver **92** are shown in Fig. 3. An actual system would Include similar circuits **90** and **92** for each of the three phases.

The minimum ON time and "DEAD" time generator **90** includes a comparator **94** which receives the PWM signal. A capacitor **95** is coupled to the comparator output by a charging path through diode **96** and a discharging path through diode **97**. When the pulse signal supplied by comparator (amplifier) **94** goes positive, capacitor **95** is charged to this value very rapidly via forward biased diode **96**. Subsequently, when the output of comparator (amplifier) **94** drops to zero, capacitor **95** discharges via diode **97**. The voltage stored in capacitor **95** is supplied to one input of comparator **98** for a Schmitt trigger comparison with the voltage on a voltage divider **99**. The parameters of the circuit are adjusted to provide a turn-on delay to ensure that there will be no "shoot through" where both power transistors of a complimentary pair are conductive at the same time. The circuit also protects against "double pulsing" where a transistor receives a turn-off pulse when it is only partially turned on. To avoid such double pulsing, the circuit delays the subsequent turn-off pulse until 2 microseconds after the turn-on pulse.

Comparator (amplifier) **98** supplies the PWM pulse train to the primary winding **40** of pulse transformer **34** via an amplifier **100** which functions as a driver for the pulse transformer. Primary winding **40** is connected to the positive supply through a resistor **102**. A diode **101** is connected across the series combination of resistor **102** and winding **40**. The purpose of the diode is to bypass inductive spikes.

The power stages connected to the secondary winding **42** of the pulse transformer **34** are shown schematically in Fig. 4. The power stages include a Schmitt trigger **110** used to reconstruct the PWM pulses, a predriver stage **112** for amplifying the PWM pulses and a common base power stage **114** which provides the driving signal for the fate of the main power transistor **30**.

Power transistor **30** is an isolated gate bipolar transistor (IGBT). Such transistors are intended to operate either in the saturated ON state of the blocked mode OFF state. The IGBT transistor is turned on by the application of a pulse to the gate and thereafter remains in the conductive state until pulsed in the opposite direction.

Schmitt trigger **110** includes a comparator **120** with a resistor **122** connected from the output to the plus input of the comparator. The plus input is also connected to ground via a resistor **124**. Secondary winding **42** of the pulse transformer is connected between the negative input of the comparator **120** and ground. A pair of Zener diodes **126, 128** is connected across secondary winding **42**. The ratio of resistors **122** and **124** can be set to have trigger points at +8V and -8V. However, the trigger points are not critical and can be in the range of 1V to 14V.

In the operation of the Schmitt trigger assume that the output of comparator **120** is initially negative and that the positive input is, therefore, set to -8V. When a positive pulse appears on the transformer secondary winding **42** and exceeds the value of +8V, the circuit triggers to the ON state. In the ON state where the comparator output is +15V, the plus input shifts to +8V. The comparator output stays positive until such time as a negative pulse appears on secondary winding **42** and exceeds - 8V. At this point, the trigger circuit reverts to the initial OFF state. In this fashion, the spike pulses passing through the pulse transformer corresponding to the leading and trailing edges of the PWM pulses are effective to reconstruct the PWM pulse.

In order to get maximum rise time, it is desirable that the pulse transformer produce spike pulses exceeding 15 V. Zener diodes **126, 128** limit the voltage appearing at the comparator input to protect the circuit components.

The predriver circuit **112** includes a PNP bipolar transistor **134** and two transistors **136** and **138** interconnected in totem-pole fashion. The base of transistor **134** is connected to the junction of resistors **130** and **132** connected between the plus supply and the output of comparator **120**. The emitter of transistor **134** is connected to the +16V supply and the collector is connected to the -16V supply via series resistors **140** and **142**. The junction of resistors **140** and **142** is connected to the bases of transistors **136** and **138**. Transistors 134, 136 and 138 increase the pulse current sufficiently to drive the MOSFET power stage which, in turn, drives the gate of IGBT transistor **30**. MOSFET transistor **144**, connected between the +16V supply and the -16V supply, generates a voltage pulse of sufficient magnitude to render transistor **30** fully conductive in the saturated state.

Power stage **114** further includes transistor **146** between the plus supply and the gate of power transistor **30**. Transistor **146** has a low collector-emitter voltage when in the conductive state and, therefore, serves to drive power transistor **30** into the hard saturation state when transistor **146** is conductive. Diode **148** connected across the emitter-base junction of transistor **146** protects against reverse bias and maintains the power transistor **30** in the blocking mode when in the OFF state.

The motor controller according to the invention is designed to operate in the range of 6000 to 20000 Hz PWM switching rates. To achieve operation at these frequencies, it is necessary to have very fast ON/OFF switching for the PWM pulses. Pulse transformer **34** can transfer pulses with a propagation time delay of less than 100 nanoseconds in either direction. The pulse transformer, Schmitt trigger and gate drive together can transfer PWM pulses in less than 700 nanoseconds for TURN-ON and less than 350 nanoseconds for TURN-OFF. With these TURN-ON and TURN-OFF rates, a PWM pulse rate of 20000 Hz can be achieved. Even higher PWM switching rates can reasonably be achieved by using faster Schmitt trigger circuits and faster predriver and driver stages. Such high speed components are within the range of existing technology, but are relatively costly at present.

The power for the motor controller is supplied by four power supplies shown schematically in Fig. 5. The four supplies include a main supply **200** which supplies power to the motor windings via the power transistor. Also included is an analog power supply **202**, a digital power supply **204**, and a floating power supply **206**, the latter being used to provide power to the individual gate drives.

Power for the controller is supplied from a three phase source of conductors U, V and W. Capacitors **240** are connected between the conductors and the chassis ground to filter out noise which would otherwise flow back onto the source.

The analog power supply **202** is connected to the source through a transformer **220** which includes a primary winding **222** connected across two of the source conductors and a center tapped secondary winding **224-225**. Secondary winding **224-225** is connected to the analog power supply **202** to produce +15V and -15V on conductor **52** (shown in Fig. 1). The center tap of winding **224-225** is connected to analog ground **53**.

Digital power supply **204** is connected to the source through a transformer **210** which includes a primary winding **212** connected across two conductors from a three-phase source. The transformer also includes a secondary winding **214** connected to the digital power supply **204**. The digital power supply produces a +5V on conductor **50** relative to a digital ground **51**.

The digital power supply is connected to supply the digital components **22** as shown in Fig. 1 and the analog power supply is connected to supply the analog components **24**. The digital power supply has a ground return path separate from the analog power supply. The two ground return paths are tied together to the supplies such as by connection **226**. Also, the digital power supply and the analog power supply are connected to the source through separate transformers **210** and **220**. By interconnecting the ground returns at the source and by

utilizing separate transformers, the two supplies are isolated from one another.

The floating power supply **206** is connected to the source via a secondary winding **216** in transformer **210**. Power supply **206** provides a full wave rectified current to a capacitor **236** in a Zener power clamp circuit and the power stored in the capacitor is supplied to a push-pull oscillator **208**.

The Zener power clamp **209** includes an NPN transistor **230** so connected that the current from the supply passes through the collector-emitter circuit of the transistor to charge capacitor **236**. A Zener diode is connected between the base of the transistor **230** and ground and current flow through the Zener is supplied by a resistor **234** connected across the collector-base of the transistor. A diode **232** is connected across the collector-emitter circuit of the transistor. The Zener diode is selected having a voltage threshold of 14.5 V. In operation, the Zener clamp circuit **209** permits current flow from supply **206** through transistor **230** while the potential across the capacitor is below 15 V. When the potential across the capacitor exceeds the Zener potential and the base-emitter voltage drop of the transistor, the transistor becomes non-conductive. As a result, the circuit clamps the potential across the capacitor at a potential slightly above the Zener potential.

Floating power supply **206** supplies power to the gate drive circuits. The supply to the gate drive circuits is preferable at a relatively high frequency to reduce the size of the isolating transformers (shown in Fig. 6). Accordingly, a push-pull open loop control circuit **208** is used which is in the nature of a push-pull oscillator operating at approximately 60 KHz. MOSFET transistors **238** and **239** are connected to the oscillator to operate in push-pull fashion such that the transistors alternately become conductive. The result is a pair of output currents designated PPL (Push-Pull Left) and PPR (Push-Pull Right) which are 60 KHz rectangular waves which become positive alternately. PPL, PPR and the V_{clamp} conductors connect to transformers which provide independent floating power to each of the six gate drives for the power transistors supplying the motor as shown in Fig. 6.

The main supply **200** includes three diodes **246-248** and three SCRs **242-244** connected to form a three phase full wave rectifying bridge. The rectified output is supplied to capacitors **250** and **251** in series with one another between +bus **56** and -bus **58**. A resistor **252** is connected across capacitor **250** and a resistor **253** is connected across capacitor **251**. The capacitors filter out the ripple from the power supply and resistors **252** and **253** provide a discharge path when the supply is turned off.

A transformer **260** is connected to supply conductors V_{clamp}, PPR and PPL to charge a capacitor **263** via a full wave bridge rectifying circuit **262**. A potential across capacitor **263** is supplied to the gate elements of SCRs **242-244** via suitable voltage dividers. Capacitor **263** provides a DC filter for removing the high frequency ripple. A soft start for the main power supply can be achieved by delaying the initial operation of push-pull controller **208**.

It is significant that the floating power supplies which power the gate drive circuits and are connected to the secondary windings of pulse transformer **34** are coupled to the source through transformer **210**. The analog power supply **202** supplies power to the electronic analog circuits **24** which are connected to the primary winding of pulse transformer **34**. The floating power supply **206** and the analog power supply **202** are connected to the source through separate transformers **210** and **220** to thereby provide isolation between circuits on opposite sides of the pulse transformer.

A separate floating supply is used with each of the gate drive circuits associated with the six IGBT power transistors. A complementary power transistor pair **30** and **32** for Phase A is shown in Fig. 6. Specifically, the collector of the upper transistor **30** is connected to the +bus and the emitter is connected to the phase A motor winding. The collector of lower power transistor **32** is connected to the phase A motor winding as well as the emitter of transistor **30**. The emitter of transistor **32** is connected to the -bus. When a pulse is applied to transistor **30** which is positive at the gate relative to the emitter, transistor **30** becomes conductive and goes into saturation effectively connecting the motor winding to the +bus. On the other hand, if a pulse is applied to the gate of the transistor **32** relative to its emitter, the transistor is driven into saturation and connects the motor winding to the -bus.

Since the drive pulse for transistor **30** is relative to the emitter which can vary from the +bus to the -bus voltages, the drive circuit and the power supply must both be floating. For the upper power transistor, PWM-upper signal for phase A is applied at a predriver **92a** which, in turn, is coupled to the primary winding **40a** of pulse transformer **34a** previously described in connection with Fig. 2. The secondary winding **42a** of the pulse transformer is connected to the Schmitt trigger and, in turn, to the gate drive in unit **268a**.

A floating supply for the upper gate drive circuit includes a transformer **270a** having a center tapped primary winding **272a**. The center tap of the primary winding is connected to V_{clamp} (see Fig. 5) and the ends of the windings are connected, respectively, to "PPL" and "PPR". One end of the secondary winding **274a** is connected via diodes **276a** and **277a** to provide +16V and -16V, respectively, to the gate drive circuit. The other end of the secondary winding provides the floating ground which is connected to the emitter of transistor **30**. Capacitors **278a** are connected between the floating ground and the diodes **276a** and **277a**. Since the gate drive and the power supply, therefore, are both transformer coupled, the gate drive and the floating power supply can both float according to the potential appearing at the emitter of transistor **30**. The capacitors **278a** filter the supplied voltage and also bypass noise. Common mode noise is rejected because of transformer

coupling of transformer **270a**. Common mode noise is further rejected by the transformer **210** coupling the floating supply to the source, as shown in Fig. 5.

A separate floating supply is provided for the lower gate drive transistor **32**. The gate drive is connected between the base and emitter of transistor **32**. A floating supply including transformer **270b** is essentially the same as that including transformer **270a**. +16V and -16V floating power is supplied to the gate drive **268b** and the floating ground is connected to the emitter of transistor **32**.

As shown in Fig. 6, the conductor **280** connecting to phase A of the motor passes through a LEM current sensor **282**. The LEM sensor is powered from the analog component supply. The output of LEM **282** is a voltage proportional to the current passing to the motor. A similar LEM sensor may be placed on each of the conductors energizing the three phase windings of the motor. The current feedback is supplied to comparators **274** and **278** in Fig. 3 via conductors **77** and **81**.

The foregoing description includes a preferred embodiment of the invention. It should be obvious to those skilled in the art that there are many variations and other embodiments contemplated within the scope of the invention.

Claims

1. A high frequency, isolation pulse transformer (34) capable of operating at a pulse rate greater than 6000 Hz, comprising a high permeability core (66, Fig. 2), a primary winding (40) wound on said core, a secondary winding (42) wound on said core, said windings being spaced apart and configured so that the capacitance between said windings is less than 5 picofarads, and said core being saturable so that only the leading and trailing edges of applied pulses pass through said transformer (34).

2. The pulse transformer of claim 1 wherein the capacitance between said windings is less than 1 picofarad.

3. The pulse transformer of claim 1 wherein the capacitance between said windings is less than 0.2 picofarad.

4. The pulse transformer of claim 1 wherein the separation between said windings is 200-300 mils.

5. The pulse transformer of claim 1 operating in the frequency range of 6 to 20 KHz.

6. The pulse transformer of claim 1 wherein said high permeability core (66) is made of a ferrite material.

7. The pulse transformer of claim 1 wherein said windings each have between 10 and 13 turns and less than 27 gauge wire.

8. The pulse transformer of claim 1 wherein said core is configured so that the flux path surrounds said windings.

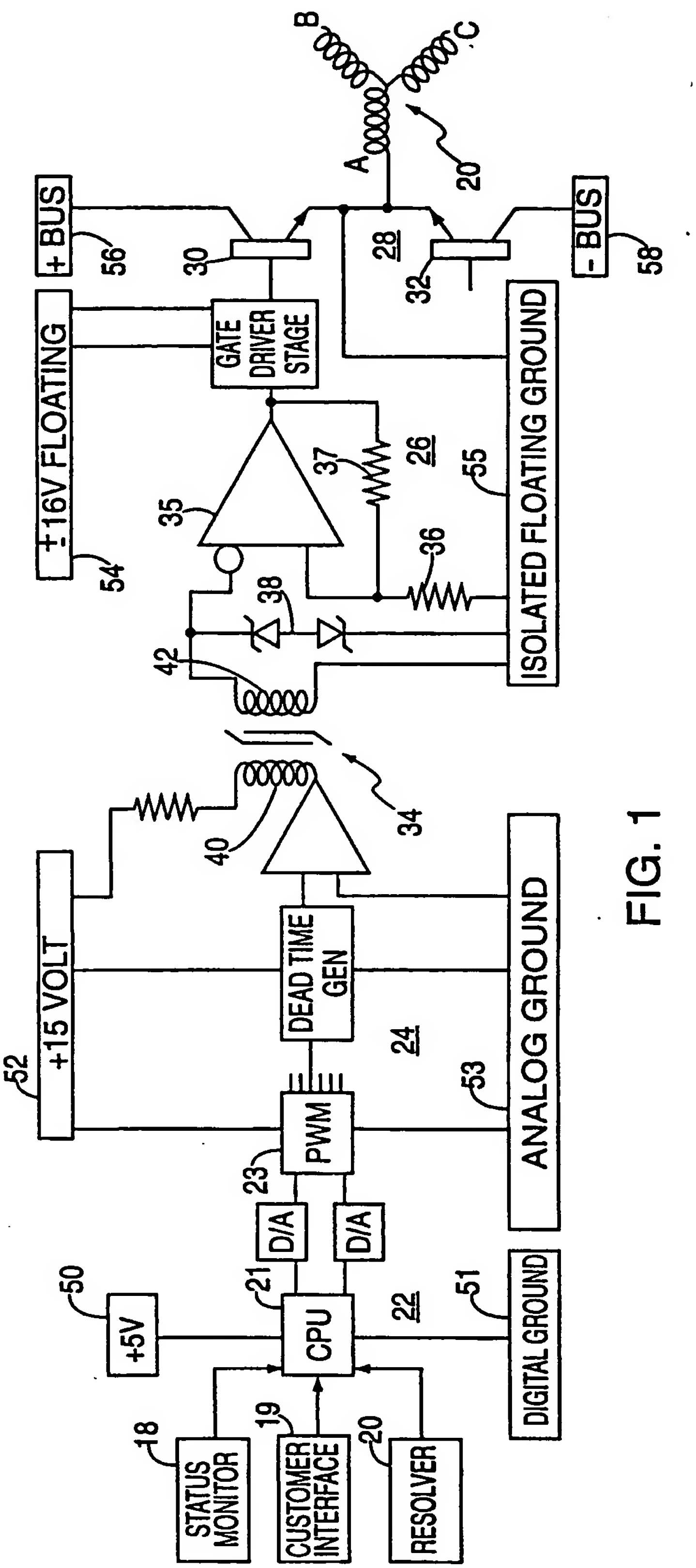

STATUS MONITOR
18
CUSTOMER INTERFACE
19
RESOLVER
20
+5V
50
CPU
21
22
DIGITAL GROUND
51
D/A
D/A
PWM
23
+15 VOLT
52
DEAD TIME GEN
24
ANALOG GROUND
53
40
34
42
38
35
36
37
26
±16V FLOATING
54
GATE DRIVER STAGE
ISOLATED FLOATING GROUND
55
+ BUS
56
30
28
32
- BUS
58
A
B
C
20

FIG. 1

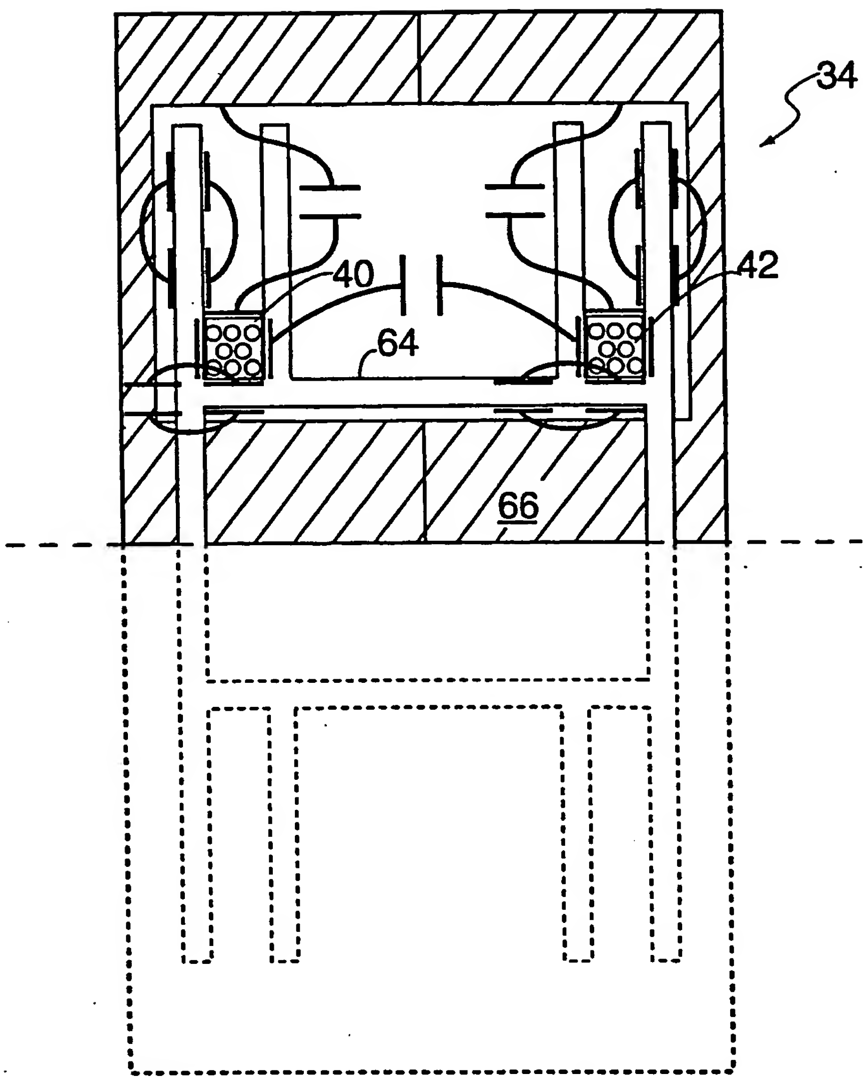


FIG. 2

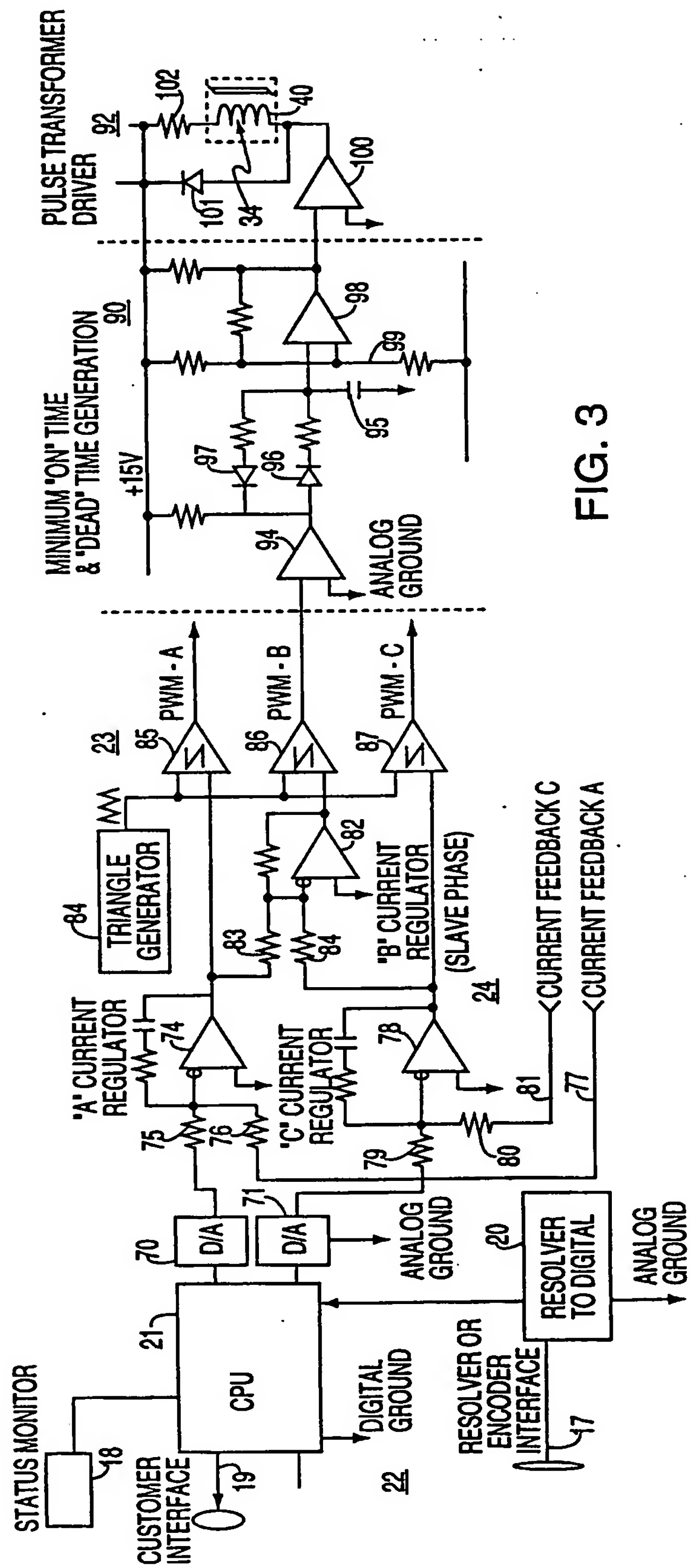
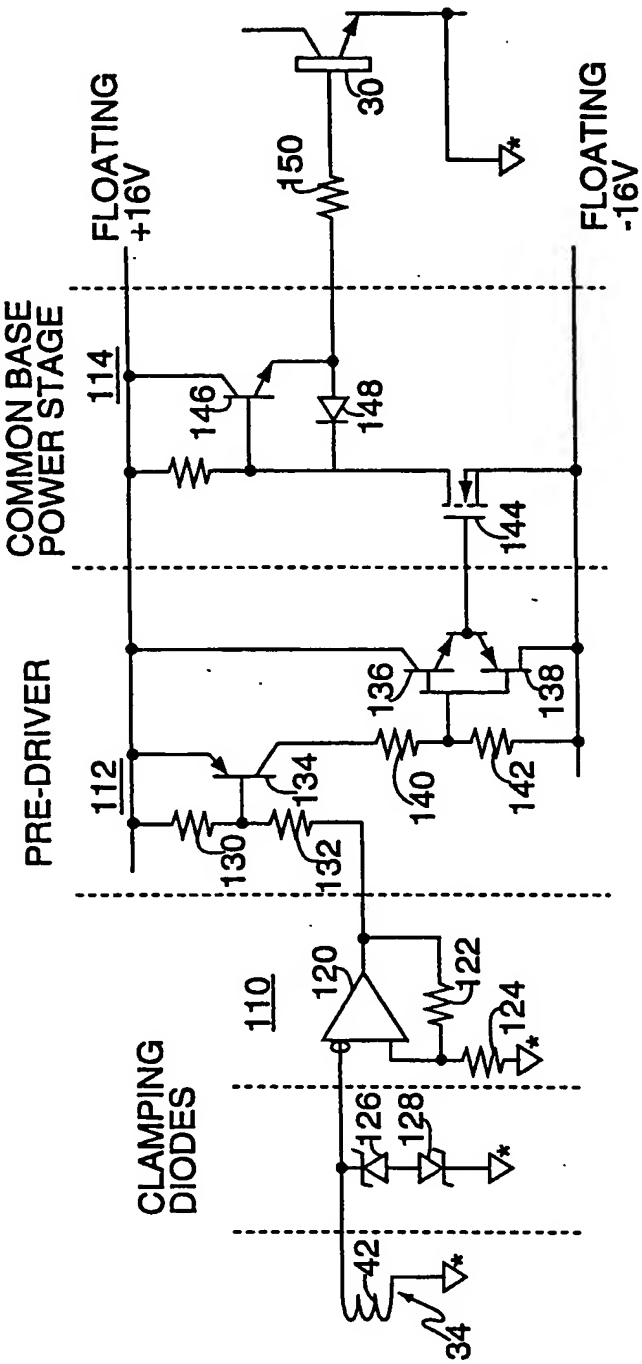

PULSE TRANSFORMER DRIVER
92
102
40
100
101
34
MINIMUM "ON" TIME & "DEAD" TIME GENERATION
90
+15V
98
99
97
96
95
94
ANALOG GROUND
PWM - A
PWM - B
PWM - C
23
85
86
87
TRIANGLE GENERATOR
84
82
83
84
"B" CURRENT REGULATOR
(SLAVE PHASE)
24
CURRENT FEEDBACK C
CURRENT FEEDBACK A
"A" CURRENT REGULATOR
74
75
76
"C" CURRENT REGULATOR
78
79
80
81
77
D/A
D/A
70
71
ANALOG GROUND
RESOLVER TO DIGITAL
20
ANALOG GROUND
CPU
21
STATUS MONITOR
18
CUSTOMER INTERFACE
19
DIGITAL GROUND
22
RESOLVER OR ENCODER INTERFACE
17

FIG. 3



* ISOLATED FLOATING GROUND

FIG. 4

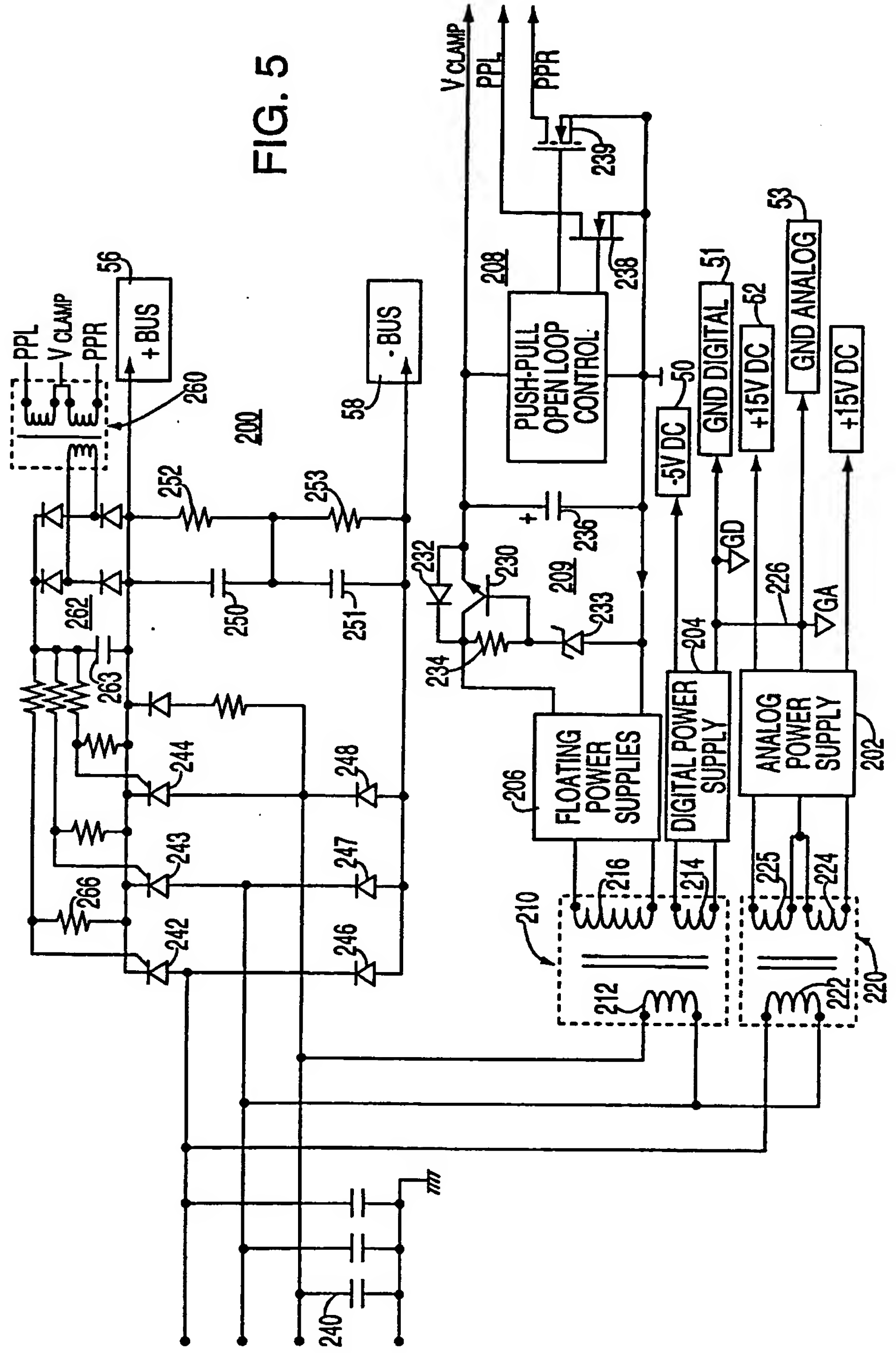

+BUS
-BUS
PPL
V CLAMP
PPR
PUSH-PULL OPEN LOOP CONTROL
FLOATING POWER SUPPLIES
DIGITAL POWER SUPPLY
ANALOG POWER SUPPLY
-5V DC
GND DIGITAL
+15V DC
GND ANALOG
+15V DC
GD
GA
200
202
204
206
208
209
210
220
240
260
262

FIG. 5

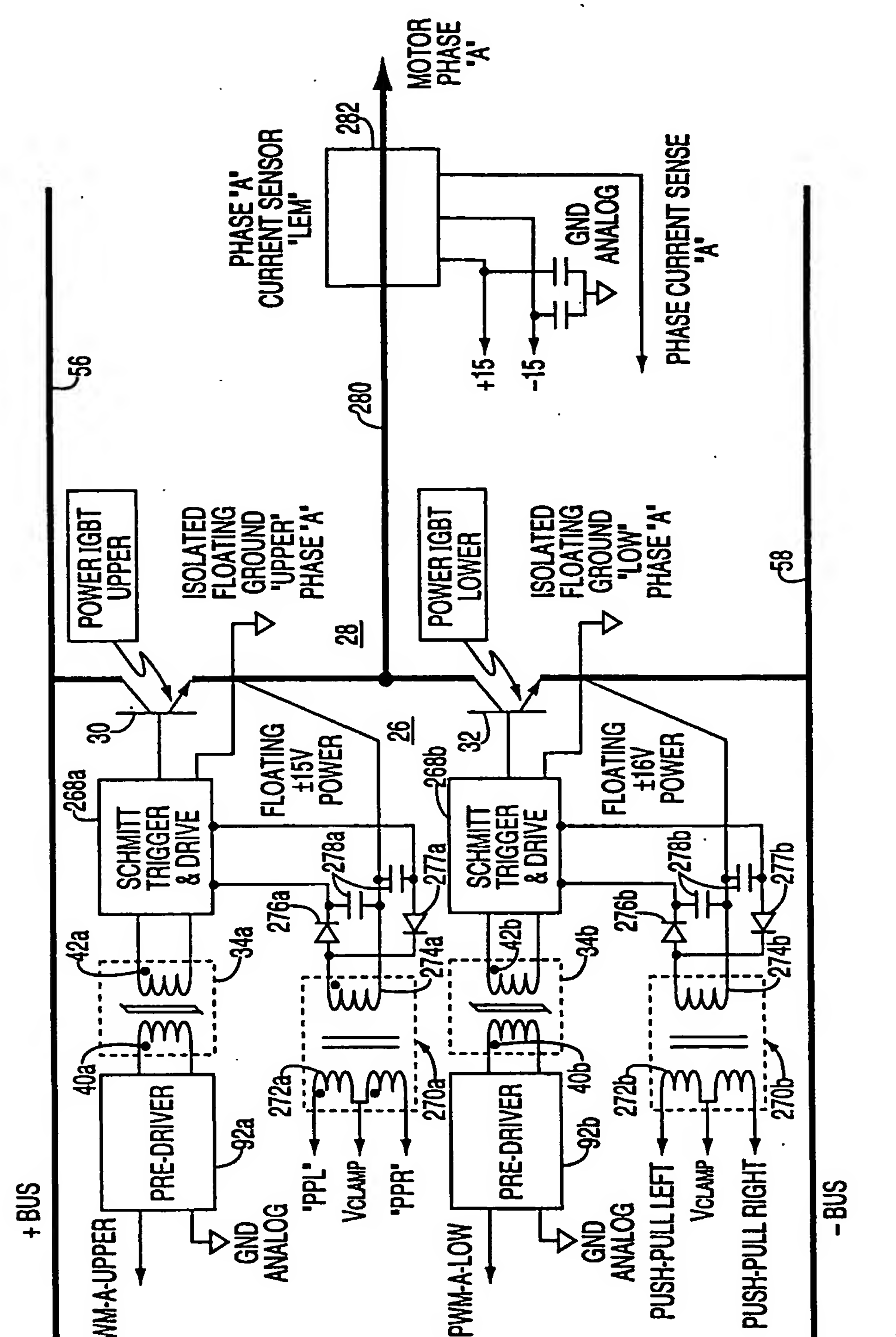

+BUS
-BUS
PWM-A-UPPER
PRE-DRIVER
GND
ANALOG
"PPL"
VCLAMP
"PPR"
SCHMITT
TRIGGER
& DRIVE
FLOATING
±15V
POWER
POWER IGBT
UPPER
ISOLATED
FLOATING
GROUND
"UPPER"
PHASE "A"
PWM-A-LOW
PRE-DRIVER
GND
ANALOG
SCHMITT
TRIGGER
& DRIVE
FLOATING
±16V
POWER
POWER IGBT
LOWER
ISOLATED
FLOATING
GROUND
"LOW"
PHASE "A"
PHASE "A"
CURRENT SENSOR
"LEM"
GND
ANALOG
+15
-15
PHASE CURRENT SENSE
"A"
MOTOR
PHASE
"A"
"PPL" PUSH-PULL LEFT
"PPR" PUSH-PULL RIGHT
56
58
28
26
30
32
280
282
268a
268b
42a
42b
34a
34b
40a
40b
92a
92b
272a
272b
270a
270b
274a
274b
276a
276b
277a
277b
278a
278b

FIG. 6